

Quadrature-Spoof Surface Plasmon Polariton Unit Cell With Circuit Characterization for Circular-Polarization Generation

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Abstract—A quadrature-spoof surface plasmon polariton (SSPP) unit cell (QS-UC) is proposed and theoretically characterized for circularly polarized (CP) wave generation. The proposed QS-UC is composed of a fishbone SSPP waveguide (FSW) and a spine quadrature mode converter (SQMC), which converts the SSPP mode into a quadrature radiating mode. An equivalent circuit is established to derive the cascaded ABCD matrix and the conditions for quadrature CP generation. Scattering parameters and dispersion characteristics are extracted from the established circuit model. The generated CP far-field wave is calculated through the quadrature-mode current distribution on the SQMC. The theoretically calculated results are also compared to the EM-simulated results, with good agreement. Based on the proposed QS-UC, a quadrature SSPP linear array (QS-LA) is developed, fabricated, and measured for the verification of CP generation. Detailed measured results show that the developed QS-LA has a wide impedance bandwidth of 5.28–11 GHz (70.3%). More importantly, the developed QS-LA shows a wide CP bandwidth covering 5.8–10 GHz (53.2%), with CP beams covering (−65°, 16°) for right-hand CP(RHCP) generation at port 1 and symmetrical range of (−16°, 65°) in left-hand CP(LHCP) generation at port two. These excellent performances verify the effectiveness of the proposed QS-UC in CP generation.

Index Terms—Circular polarization, equivalent-circuit model, quadrature mode, spoof surface plasmon polaritons (SSPPs).

I. INTRODUCTION

BY MIMICKING the behavior of surface plasmon polaritons in optics, spoof surface plasmon polaritons (SSPPs) show the same spatial confinement of electromagnetic waves

and strong slow-wave transmission properties [1]. Compared to natural-materials-designed microwave components, SSPP-based components have the advantages such as compact size, multiple resonances, high confinement of the surface wave, and so on. Owing to these attractive engineered physical characteristics, SSPPs have wide applications in microwave, millimeter wave, and terahertz wireless communication systems [2], [3].

As an effective surface-wave transmitter, SSPPs can be designed as transmission lines (TLs) [4], [5], [6], [7] for highly efficient transmission. In addition, versatile SSPP-based microwave devices, such as couplers [8], [9], power dividers [10], [11], [12], biomedical sensors [13], crossovers [14], circulators [15], filters [16], [17], [18], [19], and so on are widely researched. In these works, equivalent-circuit models and cascaded ABCD matrices are employed to analyze the SSPP unit cells and predict the relevant scattering and dispersion characteristics. In [6], an equivalent-circuit model of the proposed slow-wave SSPP TL is presented, which is then used to convert the SSPP mode into a quasi-TEM mode of microstrip lines. An SSPP interdigital coupled coupler [9] is proposed, with its extracted equivalent-circuit model to characterize the related transmission and dispersion diagram, showing much broader bandwidth than the WR-4 standard waveguide. In addition, SSPPs exhibit many successful applications in filters [16], [17], [18], [19]. In [18], flexible and stretchable lamp-strip-shaped SSPPs are etched on a substrate-integrated waveguide to design X-band filters, and a 2-D TL-based equivalent circuit model is established to analyze the proposed unit cell.

SSPPs have also been incorporated into various antenna designs [20], [21], [22], [23], [24], [25]. Periodically loaded SSPPs can be designed as an endfire antenna [20] when one terminal is open. An SSPP-based slow-wave half-mode substrate-integrated waveguide is proposed, and its series-fed circular patch antenna [21] is developed for highly efficient transmission and radiation. In [22], a sinusoidally modulated SSPP unit cell is developed to move the dispersion curve into the fast-wave region, enabling wide scanning beams are realized from the backward to broadside zone. In addition, to realize the polarization conversion from the linear SSPP mode into quadrature circularly polarized (CP) mode, SSPP-based CP antennas have been developed [23], [24], [25], [26], [27], [28] using different techniques. In [23],

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the SSPP waveguide works as a polarization converter, which is inclinedly inserted to convert the rectangular waveguide mode into quadrature CP mode. Furthermore, the quadrature CP mode can be excited by introducing parasitic circular patches beside the main SSPP waveguide in [26] and [27], and both prototypes show wide beam-steering CP performance. In these works, CP radiation is successfully generated and demonstrated through both simulation and measurement. However, rigorous circuit-level analysis for quadrature mode realization and dispersion analysis remains unexplored.

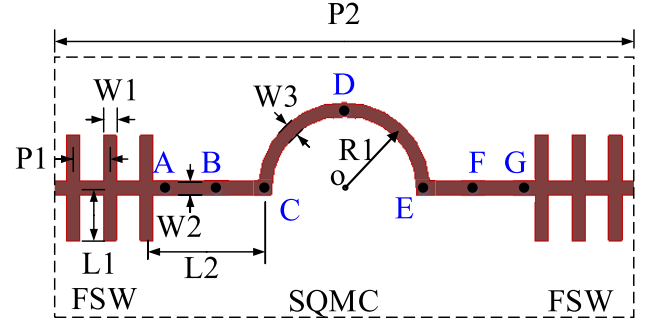
Normally, SSPPs support only a linearly polarized (LP) surface-wave mode; a quadrature CP mode is not inherent in SSPPs. How to convert the SSPP mode effectively into quadrature CP mode and illustrate it from a microwave-circuit perspective is worthy of deep research. To address this gap, this article proposes and theoretically characterizes a quadrature SSPP unit cell (QS-UC), which is composed of a fishbone SSPP waveguide (FSW) and a spine quadrature mode converter (SQMC). Based on the proposed QS-UC, the quadrature SSPP linear array (QS-LA) is then developed, fabricated, and measured for the performance verification of the CP generation, showing an excellent CP performance over the frequency bandwidth of 5.8–10 GHz, with symmetrical wide-beam range of $(-65^\circ, 16^\circ)$ or $(-16^\circ, 65^\circ)$ for right-hand CP (RHCP) or left-hand CP (LHCP) radiation. It is noted that wideband performance has also been reported in non-SSPP leaky-wave antennas, where the bandwidths can be close to 70% [31], [32]. However, these designs are based on conventional microstrip or periodic structures, and do not employ SSPP-based slowwave transmission mechanisms.

The main contributions of this work are summarized as follows.

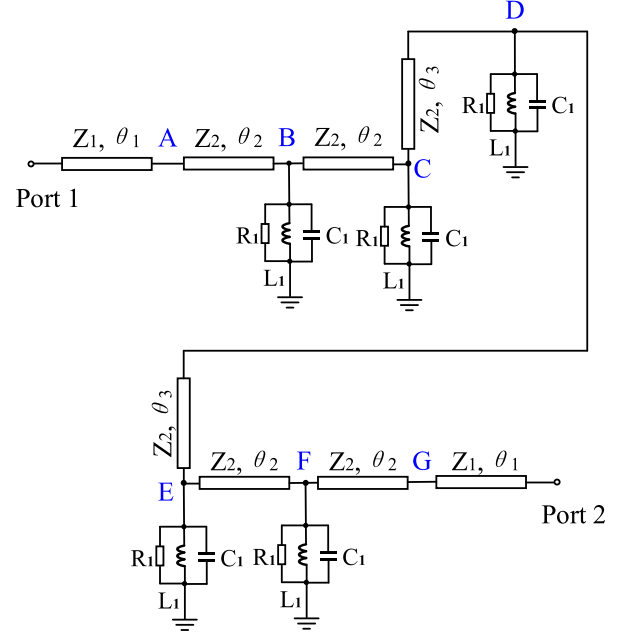
- 1) *QS-UC Architecture*: A QS-UC is proposed, consisting of the side FSW and the center SQMC. FSW provides SSPP mode transmission, while SQMC converts the SSPP mode into a quadrature radiating mode for CP generation.
- 2) *Equivalent-Circuit Model and Quadrature Mode Formation Theory*: An equivalent circuit model for QS-UC is established, and the cascaded ABCD matrix is derived. Based on the model, the conditions on SQMC for quadrature CP generation are obtained.
- 3) *Scattering and Dispersion Characterization*: From the cascaded ABCD matrix, S-parameters and dispersion characteristics are calculated and compared with the EM-simulated performances, which show good agreement in both theoretical calculation and EM simulation.
- 4) *Radiation and CP Verification*: Based on the quadrature mode current on SQMC, the radiation patterns are analyzed, calculated, and compared to EM-simulated results, showing good agreement.
- 5) A QS-LA utilizing the proposed QS-UC is developed for final verification of CP generation, which shows broadband impedance and excellent CP performance.

II. QUADRATURE SSPP UNIT CELL

In this section, a novel QS-UC architecture is proposed, with its equivalent-circuit model established to derive the



(a) Configuration of the proposed QS-UC.



(b) Equivalent circuit of the proposed QS-UC.

Fig. 1. Proposed QS-UC. (a) Configuration (detailed parameters in the figure: $P1 = 1.4$ mm, $L1 = 2$ mm, $W1 = 0.5$ mm, $L2 = 4.7$ mm, $W2 = 0.5$ mm, $R1 = 3$ mm, $W3 = 0.5$ mm, $P2 = 22$ mm) and (b) equivalent circuit.

quadrature conditions and characterize the relevant scattering and dispersion properties.

A. Circuit Characterization

The configuration of the proposed QS-UC is shown in Fig. 1(a). As shown in the figure, the unit cell includes FSWs at both ends and an SQMC at the center. The FSW is composed of multiple fishbones for SSPP mode transmission, while the SQMC is formed by the spine semicircle with certain extensions at both ends. In the structure, SQMC converts the transmitted SSPP mode from the FSW into a quadrature radiating mode.

To analyze its scattering and dispersion characteristics, the equivalent circuit is established and shown in Fig. 1(b). Key nodes of reference points A–D are defined on the structure to facilitate the circuit analysis. As shown in the circuit, FSW is modeled as an ideal TL with characteristic impedance. Z_1 and electrical length θ_1 . It is connected to the input port and output port for SSPP mode transmission. The center SQMC is modeled as the TL with the characteristic impedance Z_2 and

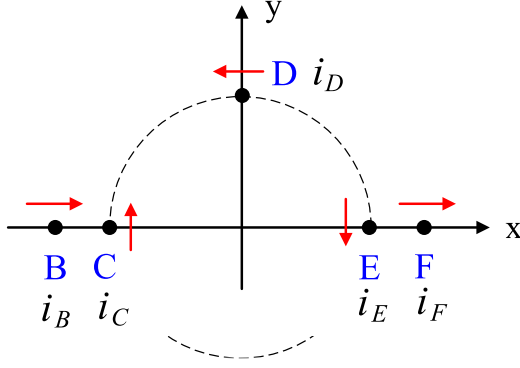


Fig. 2. Current direction defined on SQMC for quadrature mode generation.

electrical length θ_2 . In addition, considering its partial radiation characteristic, the conversion loss and introduced reactance on the SQMC are modeled as a parallel RLC circuit. In detail, the reactive components (L and C) arise from the localized storage of magnetic and electric energy on SQMC. These elements are responsible for the phase shift required for quadrature field formation. The resistive component (R) represents the radiation loss introduced by the partial conversion of the guided SSPP wave into free-space radiation and minor dissipative losses. To facilitate the analysis and without loss of generality, conductor losses on TLs are not included in the analysis.

As the FSW is directly connected to the input and output ports, its characteristic impedance can be set equal to the port impedance. Therefore, to simplify the analysis, the cascaded ABCD matrix of the SQMC from reference point A to point G is first analyzed and calculated as below

$$\begin{aligned}
 & \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{AG} \\
 &= \begin{bmatrix} \cos \theta_2 & jZ_2 \sin \theta_2 \\ jY_2 \sin \theta_2 & \cos \theta_2 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ G_A + jB_A & 1 \end{bmatrix} \\
 & \cdot \begin{bmatrix} \cos \theta_2 & jZ_2 \sin \theta_2 \\ jY_2 \sin \theta_2 & \cos \theta_2 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ G_A + jB_A & 1 \end{bmatrix} \\
 & \cdot \begin{bmatrix} \cos \theta_3 & jZ_2 \sin \theta_3 \\ jY_2 \sin \theta_3 & \cos \theta_3 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ G_A + jB_A & 1 \end{bmatrix} \\
 & \cdot \begin{bmatrix} \cos \theta_3 & jZ_2 \sin \theta_3 \\ jY_2 \sin \theta_3 & \cos \theta_3 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ G_A + jB_A & 1 \end{bmatrix} \\
 & \cdot \begin{bmatrix} \cos \theta_2 & jZ_2 \sin \theta_2 \\ jY_2 \sin \theta_2 & \cos \theta_2 \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ G_A + jB_A & 1 \end{bmatrix} \\
 & \cdot \begin{bmatrix} \cos \theta_2 & jZ_2 \sin \theta_2 \\ jY_2 \sin \theta_2 & \cos \theta_2 \end{bmatrix} \quad (1)
 \end{aligned}$$

where G_A and B_A are the conductance and susceptance of the parallel RLC circuit.

To simplify the ABCD matrix, conditions for quadrature mode conversion are analyzed first, and then they will be added into (1) for further simplification. In the circuit, supposing that quadrature mode is achieved on SQMC, the currents at the reference points B–F are defined as i_B, i_C, i_D, i_E , and i_F , and the current flowing from the left port to the right port. The directions of the current at these points are marked with arrows and shown in Fig. 2. For the reference points C–E, because

they are along the semicircle, the current directions are defined as the tangents on the semicircle. As shown in the figure, because the directions of points C and E are perpendicular to the directions of points B, D, and F, the electrical lengths between driven points C and D and points D and E must be 90° for orthogonality. Based on this, the following phase conditions must be met when there is quadrature mode current distribution on SQMC:

$$\begin{cases} |\text{phase}(i_C) - \text{phase}(i_D)| = 90^\circ \\ |\text{phase}(i_D) - \text{phase}(i_E)| = 90^\circ. \end{cases} \quad (2)$$

The remaining unknowns are the phase differences between points B and C and between points E and F. Based on the current directions in the figure and the structural symmetry, suppose that the phase differences between points B and C and between points E and F are the same as θ , that the current magnitudes at all reference points are defined as I_0 , and that the reference phase at point C is 0° , the current at these five reference points can be written as

$$\begin{cases} i_B = I_0 \exp(j\theta) \\ i_C = I_0 \exp(j0^\circ) \\ i_D = I_0 \exp(-j90^\circ) \\ i_E = I_0 \exp(-j180^\circ) \\ i_F = I_0 \exp(-j180^\circ - j\theta). \end{cases} \quad (3)$$

Then, the total current contribution along the x -axis can be summed up as

$$\begin{aligned} I_x &= I_0 e^{j\theta} - I_0 e^{-j90^\circ} + I_0 e^{-j(180^\circ + \theta)} \\ &= jI_0 (1 + 2 \sin \theta) \end{aligned} \quad (4)$$

and the total contribution of the currents along the y -axis is

$$I_y = I_0 e^{j0} - I_0 e^{-j180^\circ} = 2I_0. \quad (5)$$

From (4) and (5), it can be calculated that when

$$\theta = 30^\circ \text{ or } 150^\circ \pm 2k\pi \quad k = 0, 1, 2, \dots \quad (6)$$

the current ratio along two orthogonal directions is

$$\frac{I_x}{I_y} = j. \quad (7)$$

Therefore, a right-hand quadrature mode is achieved on the SQMC, which can be used to generate RHCP radiation wave, and the shortest electrical length of the extension is 30° .

Similarly, if the current flows from the right port to the left port, all the signs at the driven sources will be opposite, and the current ratio becomes

$$\frac{I_x}{I_y} = -j \quad (8)$$

so a left-hand quadrature mode can be achieved, which can be used to generate LHCP radiation.

After obtaining the above conditions for achieving quadrature mode on the SQMC, and substituting $\theta_3 = 90^\circ$ and

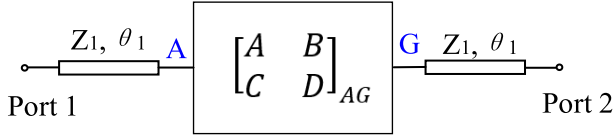


Fig. 3. Furthermore, simplified network of the full equivalent circuit.

$\theta_2 = 30^\circ$ into (1), the cascaded ABCD matrix elements can then be obtained as follows:

$$\begin{aligned}
 A_{AG} &= D_{AG} \\
 &= \frac{1}{2} - \frac{j5\sqrt{3}Y_A}{4Y_2} + \frac{15Y_A^2}{8Y_2^2} - \frac{j13\sqrt{3}Y_A^3}{16Y_2^3} \\
 &\quad + \frac{7Y_A^4}{8Y_2^4} + \frac{j\sqrt{3}Y_A^5}{Y_2^5} \\
 B_{AG} &= -\frac{j\sqrt{3}}{2Y_2} + \frac{9Y_A}{4Y_2^2} - \frac{j\sqrt{3}Y_A^2}{8Y_2^3} + \frac{29Y_A^3}{16Y_2^4} \\
 &\quad + \frac{j3\sqrt{3}Y_A^4}{8Y_2^5} - \frac{Y_A^5}{16Y_2^6} \\
 C_{AG} &= -\frac{j\sqrt{3}Y_2}{2} + \frac{Y_A}{4} - \frac{34j\sqrt{3}Y_A^2}{16Y_2} - \frac{7Y_A^3}{16Y_2^2} \\
 &\quad - \frac{10j\sqrt{3}Y_A^4}{16Y_2^3} + \frac{3Y_A^5}{16Y_2^4}
 \end{aligned} \quad (9)$$

where $Y_A = G_A + jB_A$ is the admittance of the parallel RLC.

The total circuit, including the FSW and SQMC, can now be simplified into the circuit shown in Fig. 3. The ABCD matrix of the total circuit is

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{cir} = \begin{bmatrix} \cos \theta_1 & jZ_1 \sin \theta_1 \\ jY_1 \sin \theta_1 & \cos \theta_1 \end{bmatrix} \cdot \begin{bmatrix} A_{AG} & B_{AG} \\ C_{AG} & D_{AG} \end{bmatrix} \cdot \begin{bmatrix} \cos \theta_1 & jZ_1 \sin \theta_1 \\ jY_1 \sin \theta_1 & \cos \theta_1 \end{bmatrix}. \quad (10)$$

Based on the ABCD matrix above, the impedance-matching condition and dispersion characteristic of the proposed QS-UC can then be obtained, respectively. First, to ensure that the QS-UC is matched to the adjacent cell with system reference impedance Z_1 , it can be calculated that when

$$B_{AG} = Z_1^2 C_{AG}. \quad (11)$$

QS-UC will be matched to the reference impedance Z_1 . There are a lot of possible solutions satisfying the matching condition; a simpler solution is when

$$B_A Z_2 = 0.283 \quad (12)$$

the full circuit will be matched.

Based on the matching conditions in (12), the S-parameters can be calculated by converting the ABCD matrix into an S-matrix, as shown in Fig. 4. In the calculation, the following component values are set in the circuit: $Z_2 = 2Z_0$, $R_1 = 12Z_0$, $L_1 = 1.88$ nH, $C_1 = 0.3$ pF. The EM-simulated S-parameters are also added in the figure for good comparison. As can be seen in the figure, the calculated S-parameters have good agreement with the simulated results in values

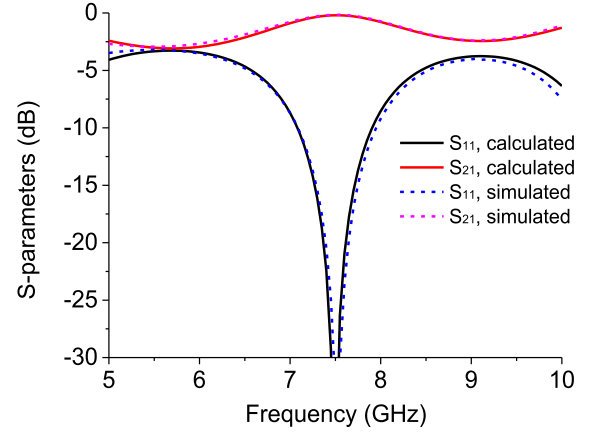


Fig. 4. Calculated S-parameters of the equivalent circuit model and simulated S-parameters of QS-UC.

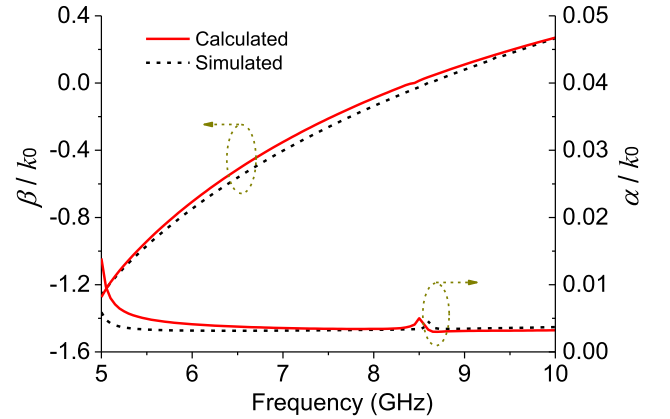


Fig. 5. Dispersion phase and attenuation constants obtained from calculation and EM simulation.

and tendency, which demonstrates the effectiveness of the developed equivalent circuit.

Its dispersion characteristic can then be obtained based on the relation between propagation constant and ABCD matrix [29]

$$\cosh(\gamma p) = \frac{A_{cir} + D_{cir}}{2} \quad (13)$$

where $\gamma = \alpha + j\beta$ is the complex propagation constant, p is the periodic length of the proposed QS-UC. Fig. 5 shows the calculated phase and attenuation constants α and β changing with the operation frequency. It can be seen that the calculated phase and attenuation constants are in good consistency with the simulated results. The small deviation between theoretical calculations and EM simulation may be mainly due to the simplifications in the parallel RLC and overlooking the coupling effect between unit cells.

Both phase and attenuation constants α and β can be adjusted by changing the related structural parameters in QS-UC. Fig. 6 shows two important parameters that affect the dispersion characteristics in the parametric study. In Fig. 6(a), it can be seen that when the radius of SQMC (R_1) changes from 2.5 to 3.5 mm, the phase constant of the unit cell gradually shifts to the left, which means a slight frequency shift to a lower frequency. For the attenuation constant, it shows stability

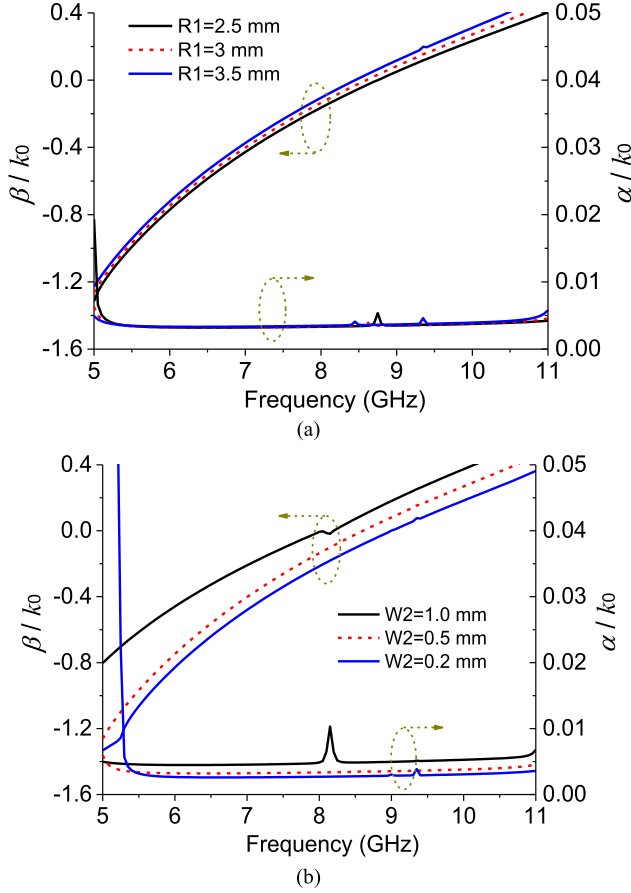


Fig. 6. Simulated normalized phase and attenuation constants changing with different (a) radius ($R1$) and (b) width ($W2$) of the proposed QS-UC.

and remains unchanged within the observed band. Therefore, we can conveniently adjust the phase constant while having very little effect on the attenuation constant.

The attenuation constant can also be adjusted, as shown in Fig. 6(b). As shown in the figure, when the width of the spine ($W2$) is reduced from 1 to 0.2 mm in QS-UC, the attenuation constant reduces accordingly. This is mainly because, as the spine width decreases, the confinement of the surface wave becomes more intense, so that more SSPP wave is transmitted, and less quadrature mode is converted. However, it should also be noticed that the change of $W2$ affects the phase constant as well; the phase constant decreases accordingly with the reduction of $W2$. Therefore, in the design of QS-UC, the attenuation constant should be first determined to achieve the desired quadrature-mode conversion rate. Once it is decided, other parameters can be further optimized to find the expected dispersion diagram suitable for the operating band.

B. Quadrature CP Generation

As the SSPP mode is converted into quadrature mode by using the proposed QS-UC, the quadrature mode will be radiated for CP wave generation. To investigate the radiation performance of the proposed QS-UC, the far-field radiation pattern is calculated by integrating the quadrature current

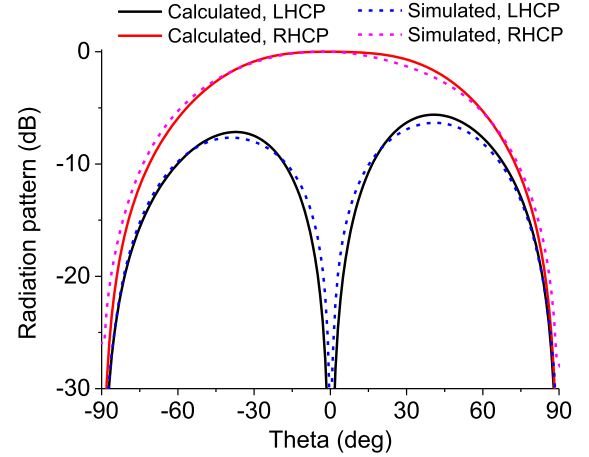


Fig. 7. Calculated normalized radiation patterns and simulated radiation patterns in the yz -plane.

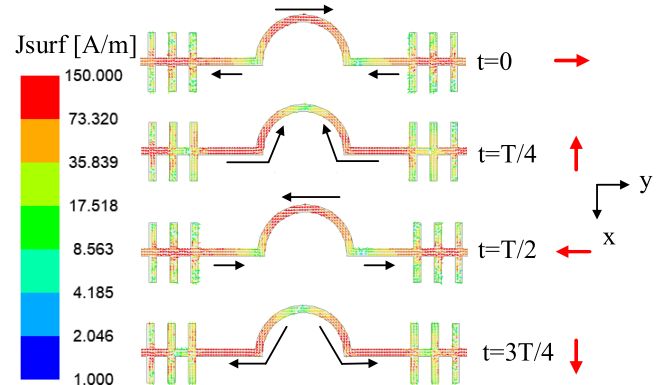


Fig. 8. Simulated current distributions on the proposed QS-UC.

distribution along the SQMC [30]

$$F(\theta, \phi) = I_0 \sin \theta \sin \phi \int_0^l e^{-\gamma \xi} e^{jk\xi} d\xi \quad (14)$$

where γ is the complex propagation constant of the current on SQMC, and k is the wavenumber in the free space.

From the above field integration along SQMC, both x - and y -components of the radiation patterns $[F_x(\theta, \phi)$ and $F_y(\theta, \phi)]$ can be obtained. Then the radiated RHCP and LHCP components can be combined as

$$F_{RHCP}(\theta, \phi) = F_x(\theta, \phi) + jF_y(\theta, \phi) \quad (15)$$

$$F_{LHCP}(\theta, \phi) = F_x(\theta, \phi) - jF_y(\theta, \phi). \quad (16)$$

Fig. 7 shows the calculated normalized CP radiation patterns in the yz -plane based on (14)–(16). The simulated results are also obtained and include for good comparison. It can be seen that a deep cross-polarization null is observed at broadside, which indicates that good RHCP radiation is realized in the broadside direction. There are also some discrepancies between the simulated and calculated radiation patterns. The slight difference is mainly due to the influence of the two FSWs beside the SQMC.

Fig. 8 shows the current distributions on the proposed QS-UC at different times of the oscillation period to indicate

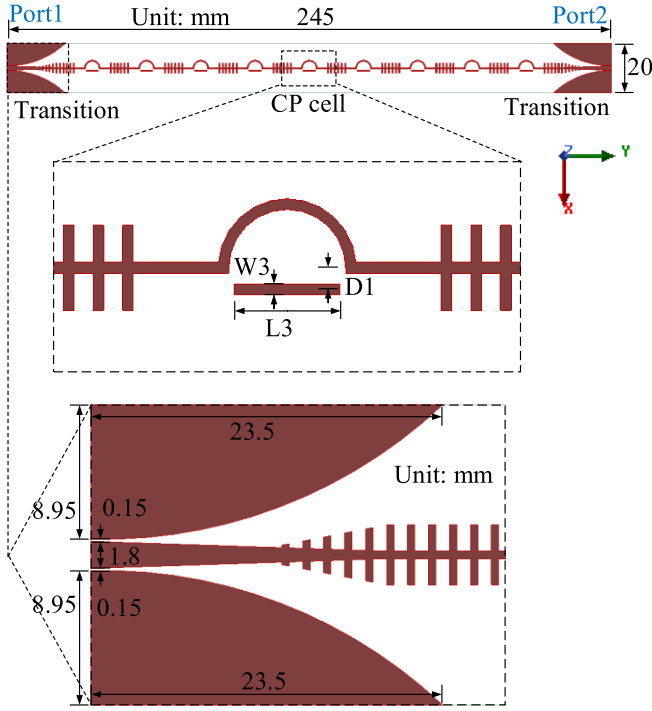


Fig. 9. Configuration of the developed QS-LA. (Detailed parameters in the figure: $L3 = 5$ mm, $W4 = 0.5$ mm, and $D1 = 1.2$ mm).

how the SSPP mode converts into quadrature mode for CP generation. In the figure, T is the oscillation period at the center working frequency. Auxiliary arrows (shown in black) are added beside the current to indicate the current direction. Red arrows are also added on the right of each figure to denote the vector summation of all the currents. It can be seen that the change of current distributions agrees well with the theoretical analysis. On the surface of SQMC, the current path is around half a wavelength along the semicircle, and some other currents will remain on the extension of the semicircle.

Furthermore, the vector summation of all the currents at each observation time is indicated by the right red arrow. At time $t = 0$, the direction of the current vector summation is along the $+y$ -axis. After a quarter period, the direction of the current vector summation is along the $-x$ -axis. The direction of the current vector summation in the remaining time becomes reversed to its counterpart. Observing the current changing with time, it can be seen that the SSPP mode is converted into quadrature mode on SQMC, and the remaining energy will continue to be transmitted to the next neighboring QS-UC. It can be inferred that based on the quadrature-mode current distribution, RHCP radiation can be realized in the boresight direction.

III. QUADRATURE SSPP LINEAR ARRAY

To verify the performance of CP generation of the proposed QS-UC, a linear array QS-LA is developed in this section.

A. Configuration

Fig. 9 shows the configuration of the developed QS-LA. As can be seen in the figure, the array consists of nine

QS-UCs for successive quadrature-mode conversion and CP radiation. In addition to the original structures designed in Section II-A, a short line strip is added below SQMC with an offset distance of $D1$ from the reference y -axis. This minor change is introduced to alleviate the minor spurious radiation caused by the structural discontinuities in QS-UC, aiming to improve the CP performance and achieve a wider bandwidth in CP bandwidth, especially at higher frequencies. This effect will be discussed in Section II-B.

The array keeps structural symmetry, so symmetrical input and output and dual-sense quadrature CP generation with the same radiation performance can be achieved when two driven ports are excited separately. For the developed QS-LA, when port 1 is excited, RHCP radiation is realized along the $+z$ direction owing to the developed quadrature mode. Whereas when port 2 is excited, LHCP radiation is realized along the $+z$ direction. The structure is designed on a single-layer PCB using the laminate of Rogers 4003C with the relative permittivity of 3.55 and a thickness of 0.813 mm. The detailed configuration of the feed transition is shown at the bottom of the figure. It converts the coplanar waveguide mode to FSW mode for convenience soldering a standard $50\ \Omega$ SMA connector. An exponential tapered ground [22] is introduced to provide gradual impedance transformation and reduce reflections over a wide frequency band.

B. Array Performance

In the QS-LA design, a short line strip is introduced to alleviate the spurious radiation caused by the structural discontinuity between the FSW and SQMC in QS-UC. This discontinuity causes the impurity in the quadrature mode, especially at higher frequencies. The line strip can be approximated as a perturbative reactive loading, which smooths the current discontinuity and reduces the phase imbalance. Therefore, it improves the quadrature conditions without altering the fundamental operation derived from the equivalent-circuit model. Fig. 10 compares the radiated axial ratio (AR) of QS-LA with and without this short line strip. As shown in Fig. 10(a), the developed QS-LA has a broad AR bandwidth for frequencies covering 5.8–11 GHz. However, without the line strip, the AR in the maximum beam direction will be deteriorated, especially in the upper frequency band.

Fig. 10(b) further investigates the far-field components rEX and rEY of the QS-LA with and without the line strip at 8 GHz. It can be seen that, compared to the rEX component, the rEY component can have more increase in its value owing to the introduction of the line strip. In addition, the phase difference between rEX and rEY is getting closer to the desired 90° , which also contributes to alleviating the quadrature CP radiation. As the line strip is along the y -axis direction, the introduction is beneficial in adjusting the magnitude of the rEY component and further improving the AR performance.

Fig. 11 shows the detailed influence of the length of the introduced line strip ($L3$). It can be seen that, when $L3$ increases from 4 to 6 mm, the radiated far-field component rEX is nearly unaffected. In contrast, the far-field of the rEY component increases gradually with the increase of $L3$. When $L3 = 6$ mm, the components rEY and rEX are

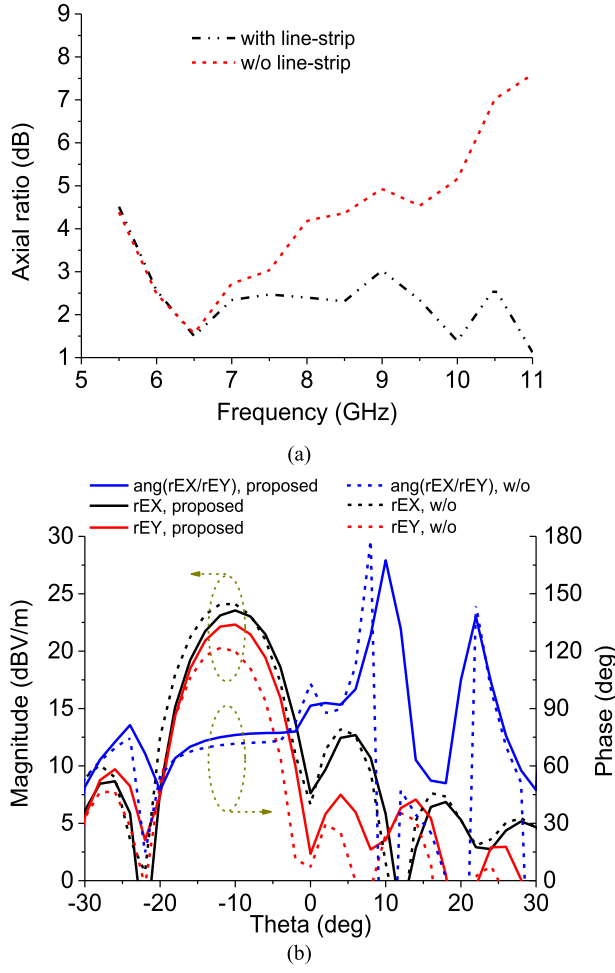


Fig. 10. Simulated (a) AR and (b) far-field components of the developed QS-LA with and without the line strip.

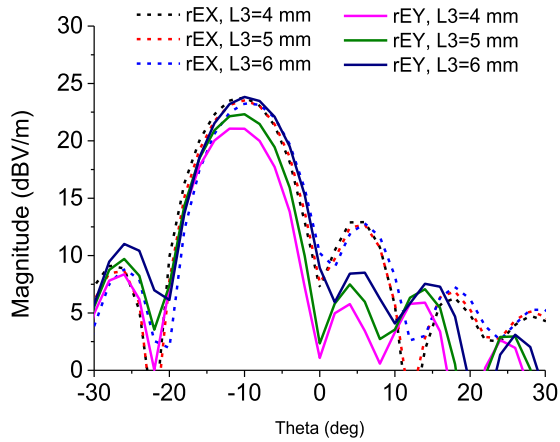


Fig. 11. Simulated far-field components of the developed QS-LA varying with different lengths of the line strip (L_3).

almost coincident in magnitude; this denotes that a high-purity quadrature CP radiation can be realized at this length. However, it should be noted that the selection of L_3 should be well balanced overall other frequencies in the bandwidth.

Fig. 12 shows the influence of the different lengths of the line-strip over the AR bandwidth of QS-LA. It can be seen in

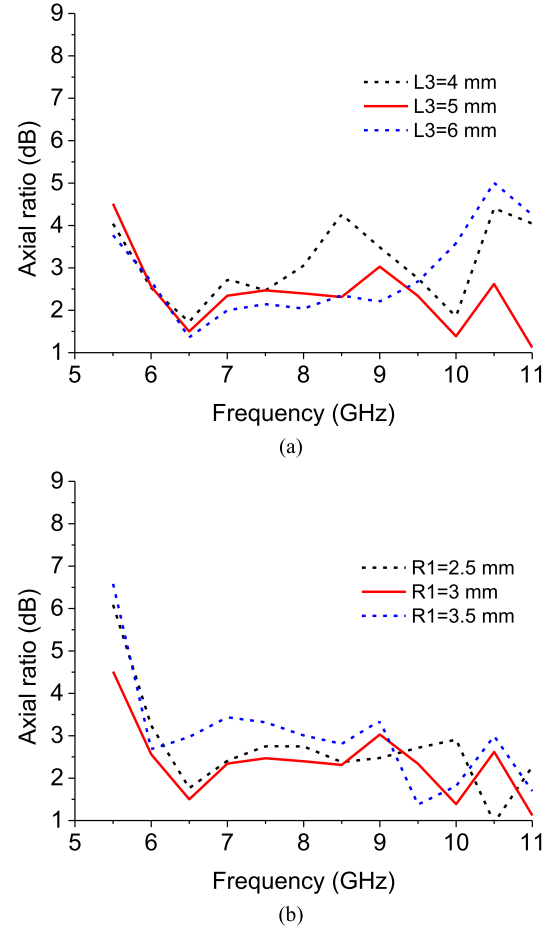


Fig. 12. Simulated AR bandwidth of the developed QS-LA varying with different (a) L_3 and (b) R_1 .

Fig. 12(a) that as the length (L_3) increases, the AR in the lower band can be gradually improved. However, the AR in the upper band will be deteriorated to some extent. This is because the line strip can couple some energy from the nearby semicircle, so the AR will be affected and improved. Therefore, L_3 should be carefully adjusted to obtain a broad frequency bandwidth. Fig. 12(b) shows the influence of the radius of the semicircle (R_1) on the AR bandwidth. It can be seen that when R_1 increases from 2.5 to 3.5 mm, the entire bandwidth has a tendency moving to the lower frequency. However, the related AR becomes deteriorated. This is because the line strip also needs to be adjusted accordingly to accommodate the bandwidth change. Overall, L_3 and R_1 are two of the most important parameters that affect the AR bandwidth and need to be finely adjusted in design.

Fig. 13 gives the simulated 3-D gain radiation patterns and E-field magnitude distribution changing with the different frequencies when the left port 1 is excited. It can be seen that the main beam shifts from the backward zone at 6 GHz to the forward zone at 11 GHz. High sidelobe appear at 11 GHz, which is due to the high-order harmonic radiation. To illustrate this phenomenon, Fig. 14 shows the simulated dispersion diagram of the SSPPs unit cell with two harmonics, β_{-1} and β_{-2} . The most relevant and significant harmonics participating

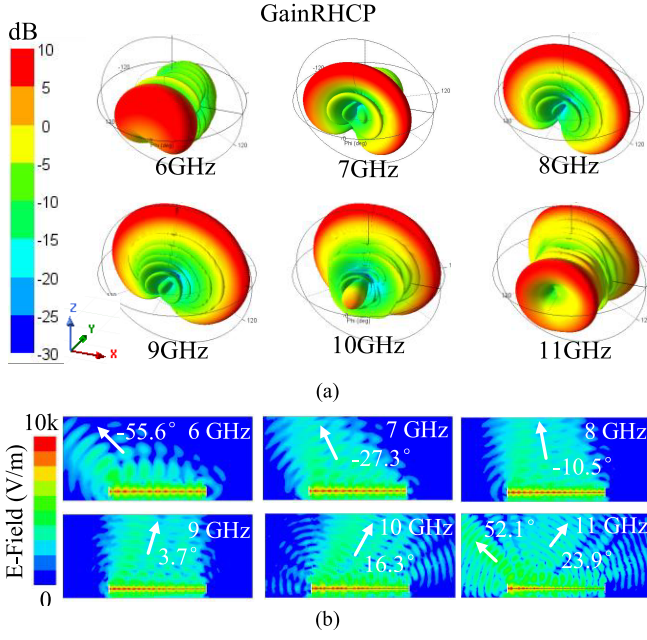


Fig. 13. (a) Scanning radiation patterns and (b) E -field distributions of the QS-LA when port 1 is excited.

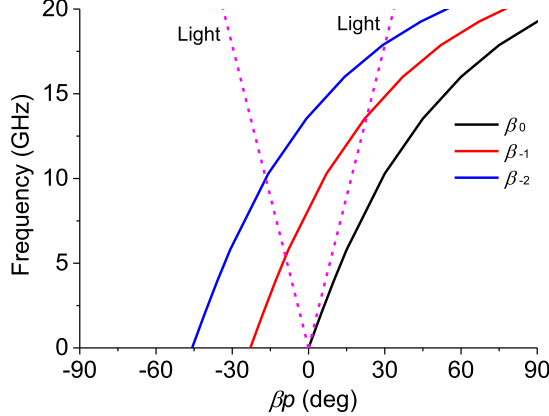


Fig. 14. Dispersion diagram of the unit cell under different harmonic modes.

in radiation are the first and second negative harmonics, as shown in the figure. The high-order harmonics are calculated as [22]

$$\beta_n = \beta_0 + \frac{2n\pi}{P} \quad (17)$$

where β_0 is the phase constant of the dominant mode on the single SSPPs, n is the order of the harmonics, and P is the period of the proposed unit cell.

As can be seen in the figure, the $n = -1$ harmonic is the fundamental radiation mode. It enters the fast-wave region at around 5.5 GHz and leaves the fast-wave region at around 13.5 GHz. The curve crosses the $\beta = 0$ axis at around 8.5 GHz. Therefore, it can be inferred that the QS-LA can have backward radiation starting from around 5.5 GHz. Due to the relatively small leakage at the lower band edge, the radiation pattern at 6 GHz appears less focused compared to the mid-band frequencies. At around 8.5 GHz, the antenna can obtain the broadside radiation. After 8.5 GHz and until 13.5 GHz, forward radiation will be realized in this frequency range.

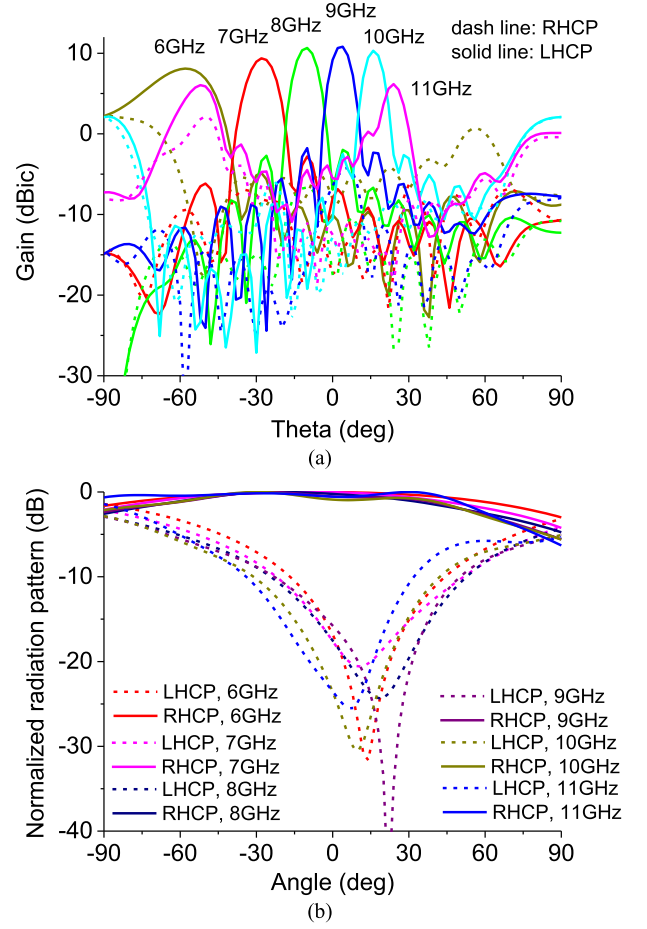


Fig. 15. (a) Simulated yz -plane scanning beams and (b) related orthogonal-plane normalized patterns changing with frequency when port 1 is excited.

This analysis agrees well with the simulated 3-D radiation patterns of QS-LA. Another should be noticed that the $n = -2$ harmonic also enters the fast-wave region as well starting from around 10 GHz. This denotes that after this frequency, two harmonics can participate in radiation. Therefore, it can be seen that at 10 and 11 GHz, sidelobes are observed in the 3-D radiation patterns. As the appearance and the participation of the radiation from the undesired higher mode $n = -2$ harmonic, the CP performance of QS-LA will be deteriorated to some extent.

Fig. 15 shows the detailed simulated radiation patterns in the yz -plane and its related beam-orthogonal plane with a step of 1 GHz. It can be seen that the beams show excellent generated CP radiation, with all the cross-polarization levels at least -15 dB lower than the copolarized main beam, ensuring the high-purity CP radiation. The gain of the main beam gradually increases from 6 to 8 GHz, but reduces thereafter from 9 to 10 GHz; this is mainly due to the appearance of $n = -2$ harmonic radiation, as discussed before. In addition, the sidelobe at the frequency of 11 GHz will have almost the same level as the main lobe, which will be excluded from the operation bandwidth in pursuit of single-beam and low-sidelobe radiation performance.

Based on the above analysis, the key parameters affecting the impedance bandwidth are studied and shown in Fig. 16, for the guidance in designing the QS-LA. Fig. 16(a) shows

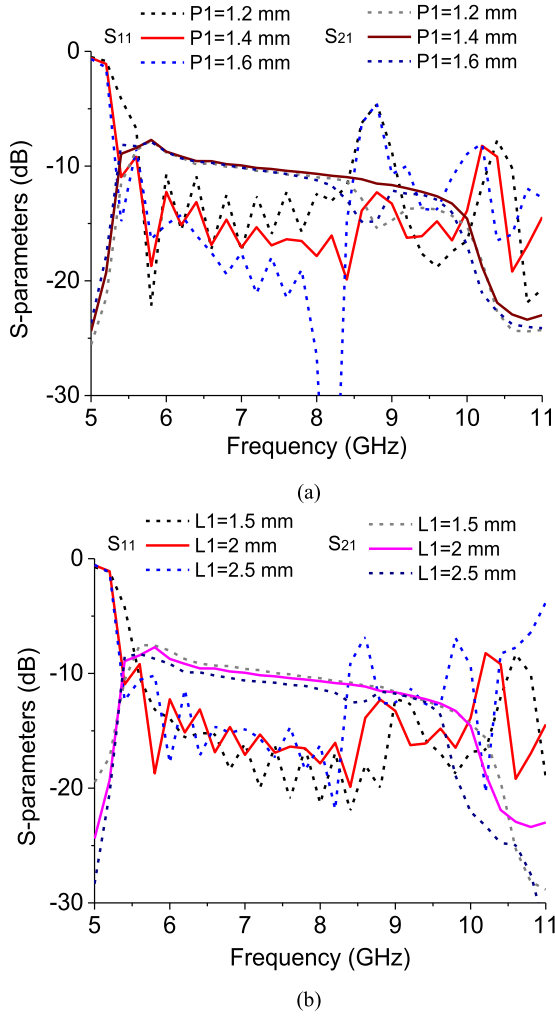


Fig. 16. Simulated S-parameters of the QS-LA varying with different lengths values of (a) P1 and (b) L1.

the effect of the period length of the SSPP (P1) on the performance of S-parameters. Note that QS-LA keeps a symmetrical structure, so only two S-parameters (S₁₁ and S₂₁) are given in the figure for brevity. It shows that a strong reflection will be created when P1 is increased or decreased, and the corresponding insertion loss will also increase a lot in the passband.

Fig. 16(b) shows the effect of another important parameter (L1), the length of the fishbone on FSW, on the impedance bandwidth. It can be seen that as L1 increases to 2.5 mm, a strong reflection is formed in the passband. It is also noticed that the bandwidth will be narrowed as L1 increases, which is mainly due to the frequency band shift of the SSPP mode. When L1 reduces to 1.5 mm, the reflection coefficient shows a narrower bandwidth. Overall, to achieve broadband quadrature CP radiation, both S-parameters and CP performance need to be taken into consideration when optimizing the relevant structure parameters.

IV. RESULTS AND DISCUSSIONS

Based on the above analyses and design, a prototype of the QS-LA was fabricated and measured for final CP generation performance verification in this section.

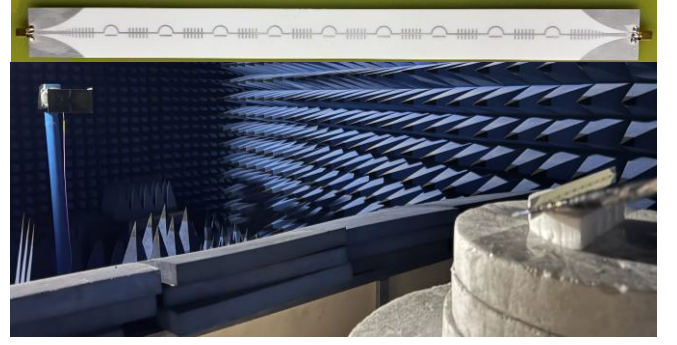


Fig. 17. Photographs of the fabricated prototype and under measurement.

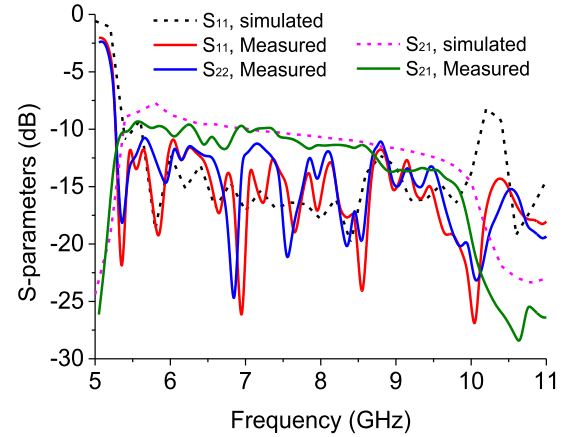


Fig. 18. Measured S-parameters of the fabricated QS-LA.

A. Verification

Fig. 17 shows the photographs of the fabricated antenna prototype. It was tested at Brunel University London and Xidian University for verification of S-parameters and radiation patterns. The measured S-parameters are shown in Fig. 18, and the simulated S-parameters are also included for good comparison. It can be seen that the fabricated antenna has a broad bandwidth covering 5.28–11 GHz with both S₁₁ and S₂₂ less than -10 dB, while the transmission coefficient is below -9.2 dB across the bandwidth. The slightly high transmission coefficient (> -10 dB) in the lower frequency band is mainly due to the low leakage radiation just moving from the slow-wave to the fast-wave zone. This can be improved by adjusting the dispersion constants illustrated in Section II-A. The measured results show good tendency compared to the simulated results. Minor discrepancies between the simulated and measured results can be observed, which could be mainly due to the fabrication errors, soldered connectors, and the connected cables.

Fig. 19 shows the measured normalized radiation patterns in the E-plane when port 1 is excited. Because of the symmetry of the developed QS-LA, only radiation patterns excited by port 1 are provided here for brevity. As can be seen, RHCP radiation is realized in the top +z region when port 1 is excited. The measured patterns show that the maximum acceptable beam-scanning range for AR < 3 dB is -65° to 16°, when the

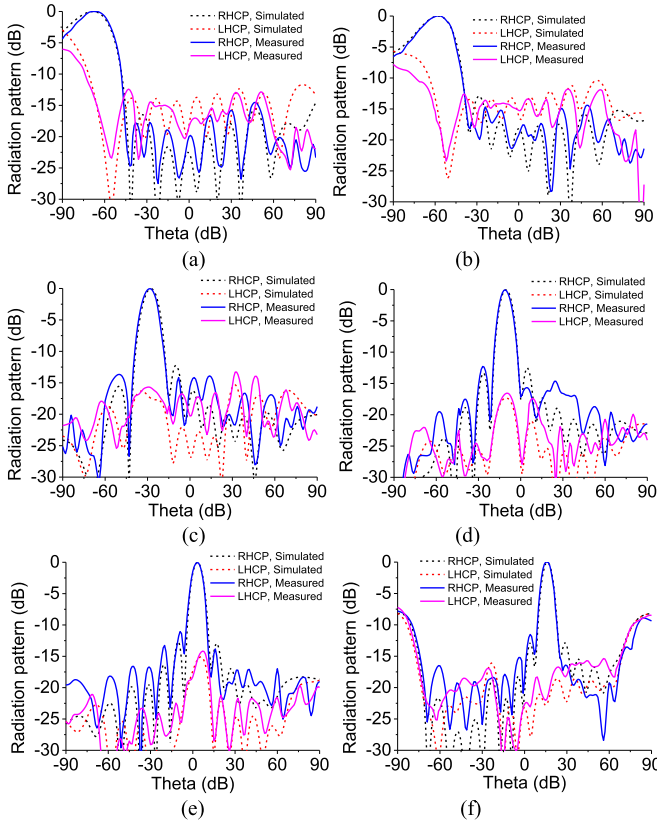


Fig. 19. Measured and simulated radiation patterns in E-plane at (a) 5.8, (b) 6, (c) 7, (d) 8, (e) 9, and (f) 10 GHz when port 1 is excited.

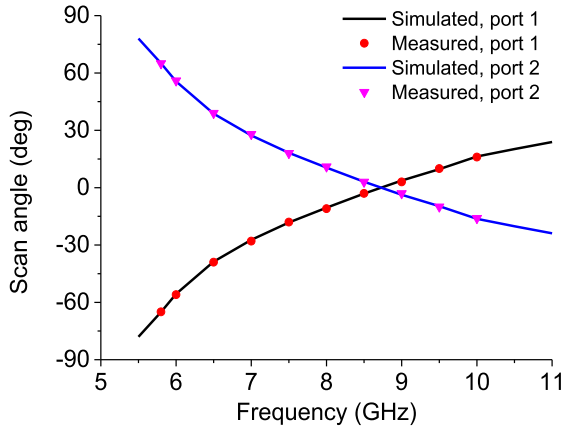


Fig. 20. Measured and simulated beam-scanning angles of the fabricated QS-LA.

frequency changes from 5.8 to 10 GHz. In addition, it is found that the beamwidth is gradually narrowed from 33° to 8° when scanning from backward to forward. It is also observed that at the upper edge frequency of 10 GHz, a large sidelobe appears in the backward zone at $\theta = -90^\circ$, which is 7.2 dB lower than the main lobe. Overall, the measured radiation patterns have good agreement with the simulated results in either the main lobe or sidelobe.

The beam-scanning capability of QS-LA is shown in Fig. 20. As can be seen, the measured CP beams show good accordance with the simulated predicted values, and

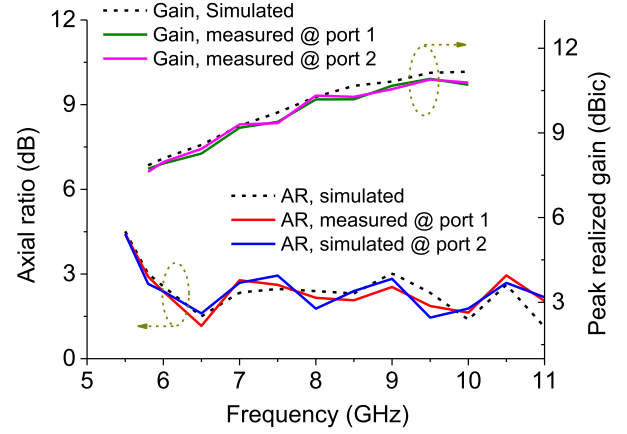


Fig. 21. Measured AR and PRG of the fabricated QS-LA.

the measured beam angles at two driven ports maintain good symmetry. The measured scanning range under $AR < 3$ dB is $(-65^\circ, 16^\circ)$ for RHCP and $(-16^\circ, 65^\circ)$ for LHCP radiation, covering the frequency bandwidth of 5.8–10 GHz. One can find that there is still some minor fluctuation between the simulated and measured beam angles, which could be from errors due to misalignment of the antenna and influence from the externally connected test cables.

The measured results of AR and peak realized gain (PRG) of the fabricated QS-LA when both port 1 and port 2 are excited are given in Fig. 21. As can be seen, the measured CP bandwidth for $AR < 3$ dB is from 5.8 to 11 GHz, which is slightly narrower than the impedance bandwidth. Compared to the simulated AR, the measured AR has a slight frequency shift to the lower frequency. The measured PRG (in consideration of the mismatch loss) is from 7.9 to 10.9 dBic for frequencies of 6–10 GHz, which is almost linearly increased with the frequency, and slightly lower than the simulated gain due to the additional insertion loss from the connector and cable. Based on the measured PRG and S-parameters, the average total efficiency (η_t) is calculated around 90% over the whole bandwidth.

B. Comparison

The main performances of the developed QS-LA are listed and compared in Table I. In the table, f_0 is the center working frequency, and λ_0 is the corresponding free-space wavelength. Designs in [23], [24], [25], [26], [27], and [28] are the recently developed SSPP-based antennas for quadrature CP radiation. As can be seen, it is very challenging to realize wideband dual CP (DCP) radiation using SSPPs. Antennas in [23], [24], and [25] can only realize single CP (SCP) radiation in either the single endfire or broadside direction. By using parasitic circular patches for CP radiation, [27] has a relatively wider scanning CP beam covering -32° to 34° with a CP bandwidth of 30.8%. The design in [28] has the narrowest unit cell width, but its scanning range covers only 10° . Many microwave-theory-driven beam-scanning radiating structures were reported in the MTT community, such as the antennas with single linear polarization (SLP) [33], [34] and dual linear polarization (DLP) [35]. As good comparison,

TABLE I
PERFORMANCE COMPARISON TO THE RECENTLY PUBLISHED WORKS

Ref.	impedance bandwidth	AR bandwidth	unit cell width	PRG (dBic)	η_t	scanning range	Polarization
[23]	8-12 GHz 40%	8.7-11.3 GHz 26%	12.9 mm $0.43\lambda_0$	16.6	/	only endfire	SCP
[24]	3.2-3.7 GHz 15%	3.2-3.6 GHz 11.7%	8.5 mm $0.1\lambda_0$	7.2	98%	only broadside	SCP
[25]	8.0-9.8 GHz 20%	8.8-9.7 GHz 9.7%	12 mm $0.37\lambda_0$	8.14	/	only endfire	SCP
[26]	10-16 GHz 46.2%	11.8-14 GHz 17.1%	7 mm $0.3\lambda_0$	10	88%	$(-19^\circ, 16^\circ)$	SCP
[27]	10.6-16 GHz 40.6%	11-15 GHz 30.8%	6.7 mm $0.29\lambda_0$	14.2	/	$(-32^\circ, 34^\circ)$ & $(-34^\circ, 32^\circ)$	DCP
[28]	8.6-11 GHz 24.5%	9.8-10.2 GHz 4%	2.93 mm $0.1\lambda_0$	9	86%	$(-46^\circ, -36^\circ)$ & $(36^\circ, 46^\circ)$	DCP
[33]	23-27 GHz 16%	0	4.5 mm $0.375\lambda_0$	14.3	21%	$(-63^\circ, 63^\circ)$	SLP
[34]	$f_0=4$ GHz	0	/	4.3	/	$(-12^\circ, 12^\circ)$	SLP
[35]	5.1-5.9 GHz 14.5%	0	38.8 mm $0.71\lambda_0$	11.6	94%	$(54^\circ, 69^\circ)$ & $(106^\circ, 128^\circ)$	DLP
This work	5.28-11 GHz 70.3%	5.8-10 GHz 53.2%	8 mm $0.21\lambda_0$	10.9	90%	$(-65^\circ, 16^\circ)$ & $(-16^\circ, 65^\circ)$	DCP

TABLE II
RECENTLY REPORTED SSPP UNIT CELL AND CIRCUIT-LEVEL WORKS

Ref.	structure type	equivalent circuit	ABCD / cascaded model	dispersion	mode / field control	polarization capability
[7]	SSPP TL	No	No	β	Slow-wave	No
[9]	SSPP coupler	Yes	Yes	β	Mode coupling	No
[14]	SSPP crossover	Yes	Yes	β	Mode coupling	No
[18]	SSPP filter	Yes	No	β	Amplitude filtering	No
[21]	SSPP fed patch	No	No	β	Series coupling	LP
This work	Quadrature SSPP UC	Yes	Yes	α, β	Quadrature field formation	CP

by using the proposed QS-UC, the developed QS-LA has a competitive impedance and CP bandwidth of 70.3% and 53.2%, respectively, with a compact width of $0.21\lambda_0$. In addition, a symmetrical wider beam coverage of $(-65^\circ, 16^\circ)$ for RHCP and $(-16^\circ, 65^\circ)$ for LHCP radiation is achieved and can be utilized for wireless dual-sense applications.

It should be noted that Table I summarizes the representative SSPP-based radiation structures primarily reported in antenna-focused venues to provide the application-level context. However, the primary contribution of this work is not the antenna design, but a UC-level theoretical characterization of SSPP structures from a microwave-circuit perspective. To further highlight the contribution from a microwave-circuit perspective, Table II compares the proposed QS-UC with other

recently reported SSPP-based TL and component-level works. It can be seen in the table that SSPP works mainly focus on transmission, filtering, or coupling; the proposed QS-UC uniquely realizes the quadrature field formation and CP generation within the circuit-level framework. The main focuses of equivalent circuit modeling, cascaded ABCD analysis, and dispersion characterization are central topics in microwave theory and techniques.

V. CONCLUSION

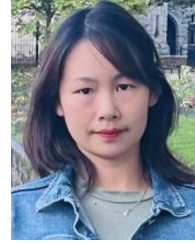
This article has presented a QS-UC and established its related theoretical equivalent-circuit model for CP generation by using SSPP structures. By combining FSW and SQMC together, the proposed QS-UC can successfully convert the SSPP mode into quadrature mode for CP generation. The equivalent-circuit model was utilized to derive its cascaded ABCD matrix and achieve the conditions for CP generation, impedance matching, and dispersion characterization. Furthermore, the generated CP radiation patterns are calculated based on the current distribution on SQMC. These characteristics and results were also validated through EM simulation. Finally, to verify the effectiveness of the CP generation of the proposed QS-UC, a QS-LA was developed, fabricated, and measured for performance evaluation. The measured results showed that the developed QS-LA has a wide impedance bandwidth of 70.3% and a CP bandwidth of 53.2%, which further proves that the proposed QS-UC and relevant theoretical circuit model are effective and feasible in utilizing SSPP structures for quadrature mode conversion and CP generation.

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